Transient simulation of lossy multiconductor interconnects

TRANSIENT SIMULATION OF LOSSY MULTICONDUCTOR INTERCONNECTS

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Abstract The transient simulation of electrically-long low-loss multiconductor interconnects is considered from a practical point of view. The importance of frequency-dependent losses in these interconnects is discussed and a simple transmission line characterization procedure allowing for such losses is proposed. The characterization obtained yields simple and efficient interconnect models, that the user can include, without programming, in any simulator accepting differential operators.

I. INTRODUCTION

Since the distortion effects of interconnects start threatening the integrity of the fast signal waveforms, the transient simulation of multiconductor transmission lines (MTLs) has become an important tool for the analysis and design of modern electric and electronic systems.

Nowadays, though the principles for an effective computation of the transient responses of interconnects are well understood (*e.g.*, see [1]), their application to the user problems can be improved. In particular, electrically-long low-loss Multiconductor Interconnects (MIs), which are widely diffused, cause particular difficulties and their practical simulation is a source of troubles. These interconnects require the use of specific TL models taking into account frequency-dependent losses, *e.g.*, the skin effect and dielectric losses [2]. Such models are not yet available in common circuit simulators (*e.g.*, SPICE) and require either lengthy coding or the time (and money) for the next generation of simulators devised for electrically-long MTLs.

In this paper the importance of the frequency dependence of losses in electrically-long low-loss interconnects is discussed and a practical approach for the transient analysis of such interconnects by standard simulators (*e.g.*, SPICE) is proposed. The approach is based on the transient matched scattering parameters of the interconnect and their numerical representation by differential operators of integer orders and ideal delays [3]. In order to emphasize the effectiveness of such an approach, simulation examples developed in SPICE and SIMULINK are also shown.

II. MODEL EQUATIONS

The problem addressed in this paper is the time-domain analysis of an electrically-long lossy $(N+1)$-conductor interconnect terminated by generic loads at both ends. We base the analysis on the voltages and currents at the ends of the interconnect conductors, v_{pq} and i_{pq}, respectively, ($p = 1, 2$ line ends, $q = 1, ..., N$ conductors, and $i_{pq} > 0$ for charges entering conductor q).

In order to obtain a transient simulation scheme of this problem which is effective also for low-loss interconnects, we describe the MTL response in terms of matched current wave variables [4], defined by

$$\mathbf{A}_p = (\mathbf{Y}\mathbf{V}_p + \mathbf{I}_p)/2 \qquad \mathbf{B}_p = (\mathbf{Y}\mathbf{V}_p - \mathbf{I}_p)/2 \ , \tag{1}$$

where $\mathbf{V}_p(s) = (V_{11}, ..., V_{1N})^T$, $\mathbf{I}_p(s) = (I_{11}, ..., I_{1N})^T$ (^{T} means matrix transposition, s is the complex frequency, and upper and lower case letters indicate Laplace transform pairs) and $\mathbf{Y}(s)$ is the line characteristic admittance matrix.

The line characteristic equations expressed in terms of current waves are:

$$\begin{array}{ll} \mathbf{I}_1 = \mathbf{Y}\mathbf{V}_1 - 2\mathbf{B}_1 & \mathbf{I}_2 = \mathbf{Y}\mathbf{V}_2 - 2\mathbf{B}_2 \\ \mathbf{B}_1 = \mathbf{H}\mathbf{A}_2 & \mathbf{B}_2 = \mathbf{H}\mathbf{A}_1 \\ \mathbf{A}_1 = \mathbf{I}_1 + \mathbf{B}_1 & \mathbf{A}_2 = \mathbf{I}_2 + \mathbf{B}_2 \ , \end{array} \tag{2}$$

where $\mathbf{H}$ is the matched transmission response of the MTL [3]. Operators $\mathbf{Y}$ and $\mathbf{H}$ are best expressed by the modal current and voltage profiles of the MTL ($\mathbf{M}_i(s)$ and $\mathbf{M}_v(s)$, respectively) as

$$\begin{array}{l} \mathbf{Y} = \mathbf{M}_i(s) diag\{Y_{mk}\} \mathbf{M}_v^{-1}(s) \\ \mathbf{H} = \mathbf{M}_i(s) diag\{H_{mk}\} \mathbf{M}_i^{-1}(s) \ , \end{array} \tag{3}$$

where $H_{mk} = \exp(-\Lambda_k \mathcal{L})$ is the transmission response of mode k ($\mathcal{L}$ line length and $\Lambda_k(s)$ mode wavenumber) and Y_{mk} is the k-th modal admittance, here defined as $Y_{mk} = s/\Lambda_k$.

III. FREQUENCY-DEPENDENT LOSSES

In this Section, we discuss the role of frequency-dependent losses thorough an example based on a 2-conductor RLC TL. For such a structure, operators $\mathbf{Y}$

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and $\mathbf{H}$ are scalar, and

$$H(s) = \exp\Big\{ - (\mathcal{L}/v)\sqrt{1 + Z_i(s)v/sZ_0}\Big\}, \qquad (4)$$

where v, $Z_i(s)$ and Z_0 are the phase velocity, the internal impedance and the lossless characteristic impedance of the TL, respectively. Fig. 1 shows the transmission step response of this interconnect ($\bar{h}(t) = \mathcal{L}_u^{-1}\{H(s)/s\}$, $\mathcal{L}_u^{-1}$ inverse Laplace Transform operator), for $v = 2 \times 10^8$ m/s, $Z_0 = 50\ \Omega$ and different internal impedance models. The time scale of Fig. 1 starts at the line delay $\tau = \mathcal{L}/v$ and is logarithmic to show both the early- and the long-time evolution of the responses. The solid curve of Fig. 1 stems from the Holt's internal impedance $Z_i(s) = R_{dc} + R_0\sqrt{2s}$, with the *dc* and skin parameters set to $R_{dc} = 30\ \Omega$/m and $R_0 = 8.6 \times 10^{-4}\ \Omega/\mathrm{m}\sqrt{\mathrm{Hz}}$, which are values typical of low-loss MCM interconnects. In this curve, the rise time is controlled by the skin effect and is short but finite, whereas the long-time tail, beyond 1 ns, is controlled by the R_{dc} term. The dashed curve stems from the frequency-independent internal impedance $Z_i(s) = R_{dc}$. Such a curve has the correct long-time evolution but zero rise-time [2]. Finally, the dotted curve corresponds to the frequency-independent model $Z_i(s) = R_{dceq}$, where R_{dceq} is an equivalent *dc* resistance, increased to account for the value of $Z_i(s)$ in the skin bandwidth.

This example points out that, in low-loss TL, losses can affect only signals with very fast rise-times, experiencing frequency-growing losses. Frequency-independent loss models miss the important initial part of the TL characteristics and are, therefore, useless to simulate the behavior of low-losses structures.

The computation of the curves of Fig. 1 is accomplished by a numerical algorithm for the inversion of the Laplace Transform [4]. Such an algorithm operates on the scattering responses of both 2-conductor and multiconductor TLs without the well known difficulties of the inverse FFT. It offers a controllable error, handles realistic frequency dependent p.u.l. parameters and yields accurate results also in the critical case of low-loss interconnects. This algorithm is our tool to generate reference transient characteristics, which are useful for parametric studies, validations as well as for the generation of computational TL models.

IV. LINE MODELS

In order to compute the transient response of a network containing a lossy MI, the interconnect characteristics $\mathbf{Y}$ and $\mathbf{H}$, which correspond to dynamic operators (the last with delays), must be represented in a form suitable to time-domain integration. The most effective representation is composed of differential operators of integer order (plus time delays for $\mathbf{H}$), because they offer high

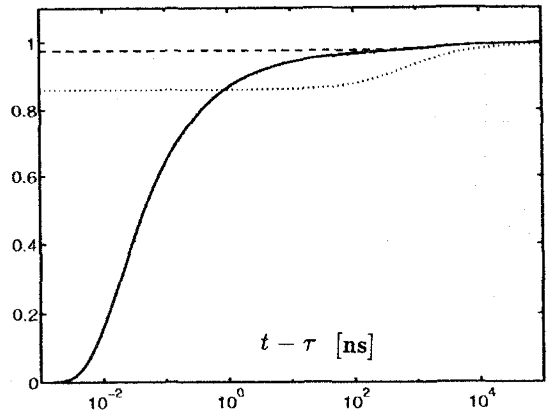

Figure 1: *Transmission step responses of a 2-conductor RLC TL for different internal impedance models (see Sec. III). This comparison highlights the importance of the frequency dependence of losses in low-loss interconnects.*

numerical efficiency and are directly accepted by many commercial simulation packages, thereby avoiding any C-level programming.

Several methods are available to convert line frequency characteristics in time-domain differential operators. The transient characteristics of a generic electrically-long multiconductor structure, however, can be very complicate and the problem makes sense only when the structures considered are properly limited. In this paper, we concentrate on the important class of low-loss interconnects with weakly inhomogeneous dielectric and nearly diagonal internal impedance matrices (*e.g.*, planar multistrip structures or wire tapes and bundles in air).

The characteristics of such structures can be effectively expressed through the approximate modal decomposition defined by $\mathbf{M}_{va} = \mathbf{M}_v(\infty)$ and $\mathbf{M}_{ia} = \mathbf{M}_i(\infty)$. In order to obtain differential operators reproducing these approximate characteristics, we rely on the approach proposed in [5], which amounts to compute the modal SRs $\bar{y}_{mk} = \mathcal{L}_u^{-1}\{Y_{mk}/s\}$ and $\bar{h}_{mk} = \mathcal{L}_u^{-1}\{H_{mk}/s\}$ and to fit them with sums of exponential functions. We compute the SRs as explained in Sec. III and fit them by means of a standard least square algorithm [3]. The outcome of the fitting step are a set of time constants and coefficients defining differential-difference operators to be inserted in any simulation environment accepting them.

The error caused by the constant modal profiles is verified by comparing the physical SRs of the exact characteristics ($\bar{\mathbf{y}} = \mathcal{L}_u^{-1}\{\mathbf{Y}/s\}$ and $\bar{\mathbf{h}} = \mathcal{L}_u^{-1}\{\mathbf{H}/s\}$) and those arising from the approximate modal decomposition. We find that, when the maximum modal delay, τ_M, and the maximum elements of the p.u.l. inductance and *dc* resistance matrices, L_M and R_M, respectively, satisfy $\tau_M \frac{R_M}{L_M} < 0.1$, the error is negligible and the comparison can be skipped.

The time-domain approach to differential representation has interesting advantages for the problem considered. It allows to predict the error of the approximation in the time-domain, and its effect on the transient simulation. This helps in minimizing the order of differential characteristics by fitting only the parts of the responses relevant to the simulation of interest. On the other hand, the numerical cost of the exponential fitting is of little concern, because the order of differential operators sought is low, usually between 2 and 6 for fittings over time intervals up to 4 decades long. Besides, the time-domain approach yields the differential operators in diagonal form, which is a numerical advantage, and handles the frequency dependence of the p.u.l. parameters without any specific care.

V. EXAMPLES

In order to show the features of our approach, in this Section, we develop examples of line modeling and transient simulation.

For a modeling example, we consider a multiconductor microstrip structure composed of four identical uniformly-spaced lands over a dielectric substrate and a ground plane (*i.e.*, $N = 4$). The structure parameter values are $\mathcal{L}$ = 10 cm, substrate height 60 μm, land-width 60 μm, -spacing 60 μm, -thickness t = 10 μm, substrate relative permettivity 9, and strip conductivity 5.6 $\times 10^7$ m$^{-1}\Omega^{-1}$. These parameter values are representative of a symmetric 4-land MCM interconnect, which is an example of structure with moderate losses and appreciable transverse inhomogeneity. The structure is modeled by differential operators generated from the approximate modal decomposition, and the result obtained for a set of physical transmission SRs is shown in Fig. 2. The aim of this Figure is to highlight, in this edge case ($\tau_M \frac{R_M}{L_M} = 0.06$) the accuracy of the model obtained by the approximate modal decomposition and exponential fitting. In fact, Fig. 2 shows the physical SRs $\bar{h}_{q2}$ and their representation by constant modal profiles and exponential fitting of $\bar{h}_{mk}$ with 3 to 5 natural frequencies. The reference and the approximate curve can be hardly distinguished. In the following, we show the use of our differential models to simulate the transient behavior of networks containing interconnects. The examples are aiming at showing how the analysis of realistic interconnects can be easily included in standard simulators by using the proposed technique.

The inclusion of the MTL model in circuit simulators is based on the Branin's equivalent circuit and on the rep-

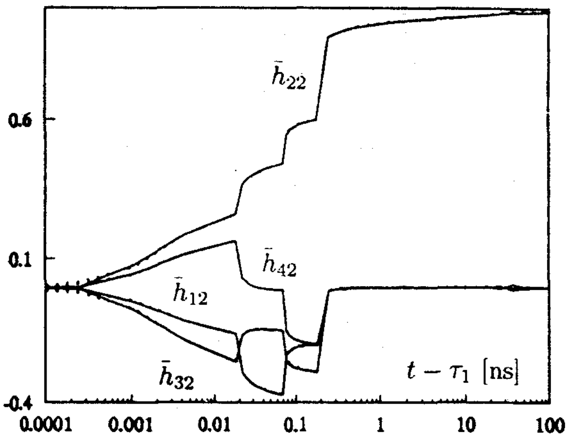

Figure 2: *Reference (solid) and approximate (dotted) transmission step responses $\bar{h}_{q2}$ (see text) for the MCM interconnect of Sec. V.*

resentation of its elements by the differential-difference operators [3]. We developed implementations of this model in different SPICE simulators. The network of Fig. 3 is used to demonstrate the operation of the PSpice version of our model. Such a network is composed of the microstrip structure modeled in the previous example driven and loaded by ECL inverters. The simulated transient is caused by a logical HIGH pulse applied by the gate driving land no. 3, while the others remain in the logical LOW state. The HIGH pulse starts at 0.1 ns, lasts 1 ns and has typical rise and fall times of 100 ps. Fig. 4 shows some of the voltage waveforms computed

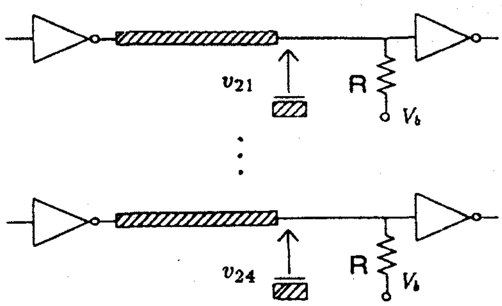

Figure 3: *Networks for the PSpice simulation example of Sec. V. Inverter devices are modeled as ECL gates, with $V_b = -2$ V and $R = 300$ Ω.*

by our PSpice model for this problem. The signal distortion for simple transmission of this case is appreciable, and arises both from the spreading of modal velocities and from the frequency-dependent losses.

The run times of our lossy models appear comparable to that of the widely used ideal lossless model, demonstrating the numerical efficiency of the approach. Besides, the lossy models shows stability properties even better than expected, offering troubleless runs also in those cases where the lossless model leads to integration problems.

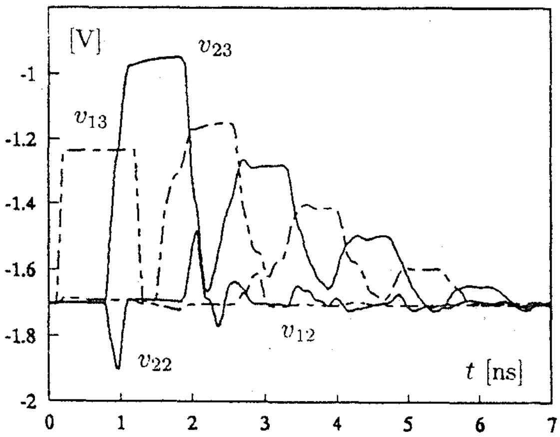

Figure 4: *Some of the end voltage waveforms in the network of Fig. 3. The response is computed for differential characteristics implemented in PSpice.*

The sequential simulators, which operate by solving the system dynamic equations via recursion, are particularly interesting for the simulation of networks with MIs, because they exploit the decoupling introduced by interconnect delays. The inclusion of the MI model in sequential simulators is strictly similar to the inclusion in circuit simulators, but relies on a block diagram interpretation of (2) [3]. We develop a sequential implementation in SIMULINK, and, in order to verify the advantage of the sequential simulation approach in highly interconnected networks, we simulate the transient behavior of the network of Fig. 5. Such a network contains simple lumped parts and 3-conductor interconnects composed of two circular wires lying over a ground plane. The parameter values for this example are $\mathcal{L} = 6$ m, $N = 2$, wire-diameter 10 mils, -separation 30 mils and -height over the ground plane 15 mils.

Some of the voltage waveforms computed by this model for the network of Fig. 5 are shown in Fig. 6. The large difference between the line delay and the pulse duration (or the line transmission rise time) requires the transient solution in a large number of time points. This kind of transient problem is a challenge for most conventional line characterization and simulation approaches. The SIMULINK model, on the contrary, solves the problem of Fig. 5 without any trouble and much quickly than the PSpice model, confirming the the expected advantage.

VI. CONCLUSIONS

We discuss the importance of frequency-dependent losses in electrically-long low-loss interconnects and propose a practical method to create differential models for these structures. The models can be developed at the level of accuracy required by any given simulation problem and can be inserted either in circuit simulators (useful for mainly lumped networks) or in sequential simulators (adequate for highly interconnected networks). In this way, realistic interconnect models with high efficiency are obtained at very low cost and without any in-depth knowledge of the simulator operation.

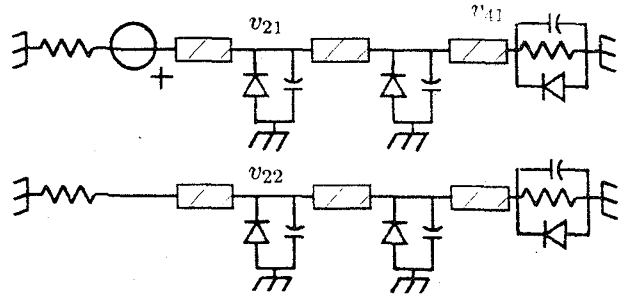

Figure 5: *Network for the SIMULINK simulation example of Sec. V. The source is a trapezoidal pulse of 1 V starting at* $t = 100$ *ps with 100 ps rise time. The other parameter values are* $R = 100\ \Omega$ *and* $C = 1$ *pF.*

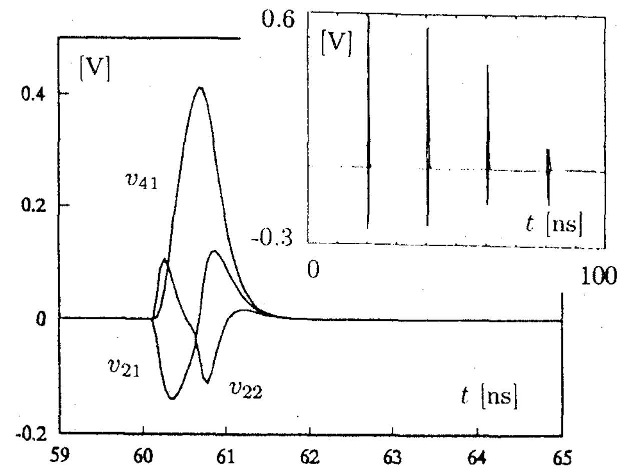

Figure 6: *Voltage waveforms at the line ends of the network of Fig. 5 (insert) and some of them around* $t = 60$ *ns (main part).*

REFERENCES

[1] D. B. Kuznetsov *et al.* , *IEEE Trans. Circuits and Syst. I*, vol. 43, pp. 110–121, Feb. 1996.

[2] A. Deutsch *et al.* , *IBM J. Res. Develop.*, vol. 34, pp. 601–614, July 1990.

[3] I. Maio *et al.* , *Proc. of EMC'96 ROMA*, pp. 678–683, Sept. 1996.

[4] I. Maio *et al.* , *Proc. of EMC Symposium, Zurich, Switzerland*, pp. 197–202, Mar. 1995.

[5] A. Semlyen *et al.* , *IEEE Trans. Power Appar. and Syst.*, vol. PAS94, pp. 561–571, Mar. 1975.